An Overview of Microwave Bandpass Filters with Capacitive Coupled Resonators

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Abstract – This paper presents a summary of several bandpass filters all consisting of resonators coupled only by capacitive coupling. The basic filter's configuration with asymmetric inductors is explained in detail as well as its modifications with asymmetric capacitors. The filter's design process is undemanding due to the filter's simple configuration, while its compact size makes it suitable for integration. Six filters with central frequencies ranging from UHF to X band were designed, realized, and tested with excellent agreement between the predicted and measured results.

Keywords – Capacitive coupling, lattice structure, bandpass filters, microstrip, printed filters.

I. INTRODUCTION

Band pass filters with low insertion losses are essential components of all kinds of modern communication devices because of both their band selecting and image and spurious rejecting capability. The latest achievements in microwave monolithic integrated circuits and high temperature superconducting technology have additionally stimulated the development of various planar filters. Because of that, almost every issue of contemporary RF and microwave scientific magazines contain several papers featuring various new configurations and design propositions for printed filters.

The basic elements of all filters presented in this paper are resonators that consist of an inductance terminated at both ends with a grounded capacitive load. The same or similar basic structure shares many forms and types of resonators appearing in numerous scientific papers under various names such as: stepped impedance resonators, hairpin resonators, resonators with interdigital capacitors, open-loop slow wave resonators, as well as many sorts of ring and half-ring resonators [1-6].



Fig. 1. Coupling methods for adjacent resonators [7]

Siniša P. Jovanović is with the IMTEL Komunikacije, Blvd. Mihajla Pupina 165B, 11070 Belgrade, Serbia, e-mail: siki@insimtel.com Resonators within a filter can be mutually coupled by capacitive (electric) coupling (Fig. 1a), inductive (magnetic) coupling (Fig. 1b) or mixed coupling (Fig. 1c) [7]. Higherorder filters usually employ all these coupling types to achieve desired characteristics. However, as shown in [8-16] it is possible to realize filters made with only capacitive coupling resonators which occupy a significantly smaller area compared to filters with inductive or mixed coupling.

II. FILTER'S CONFIGURATION

Fig. 2 shows a basic electric scheme of a filter that consists of four identical resonators (Q1 to Q4) electrically coupled by capacitors C_r . Although the scheme has only four variables, C_p , C_r , L1, and L2, by varying their values it is possible to obtain a band pass filter at a desired central frequency. This configuration is especially suitable for the printed filters with passband widths ranging from 4% to 20%, where the highest realizable Q factor of printed resonators determines the lower limit, while the upper limit depends on the highest practically realizable L1/L2 ratio (for the configuration from Fig. 2).



Fig.2. Basic electrical scheme of a filter with four electrically coupled resonators [8]





As can be seen from Fig. 2, input and output ports divide the filter into two equal, asymmetrical parts forming an anti-parallel configuration (illustrated on Fig. 3c). In that manner a lattice filter structure is formed, which is essential for its electrical property, as illustrated on Fig. 4 showing S_{21} frequency characteristics of a single asymmetrical network (a) compared with characteristics of parallel (b) and antiparallel (c) connections of the same network. Fig. 4 shows an extremely narrow bandwidth in case of a single network as well as for a parallel network connection, while only an anti-parallel connection exhibits transmission characteristics applicable for the printed filters realisation.



Fig.4. S₂₁ frequency characteristics for (a) single asymmetrical network N (gray line); (b) two parallel N networks (dotted black line); (c) two anti-parallel N networks (black line)

Although the lattice filter structures are frequently used for some types of IIR digital filters, they are hardly ever employed for analogue filter design because they require more components than the conventional ladder filter structures. However, because of its central symmetry and convenient configuration the lattice filter structure from Fig. 2 is easier for the printed filters realisation than the equivalent ladder filter structure, in spite of having an electrical scheme with more components. Similar filter characteristics can be achieved with a configuration formed by only three slightly modified resonators and with an increased value of the coupling capacitor. However, in that case the resonators would not be identical nor would the whole structure be symmetrical, so therefore it would be more sensitive to fabrication errors and less suitable for practical realisation.

With certain approximation, the central angular frequency of the filter from Fig. 2 could be expressed as:

$$\omega_{c} = \frac{1}{\sqrt{\frac{(L_{1} + L_{2})}{2} \left(C_{p} + \frac{C_{p}C_{r}}{C_{p} + C_{r}}\right)}}$$
(1)

while the angular frequency of the two zeros of the S_{21} transmission characteristic are very closely given with:

$$\omega_{z1} = \frac{1}{\sqrt{L_1 \left(C_p + \frac{C_p C_r}{C_p + C_r}\right)}}$$
(2)
$$\omega_{z2} = \frac{1}{\sqrt{L_1 \left(C_p - \frac{C_p C_r}{C_p + C_r}\right)}}$$
(3)

$$_{2} = \frac{1}{\sqrt{L_{2} \left(C_{p} + \frac{C_{p}C_{r}}{C_{p} + C_{r}}\right)}}$$
(3)

The characteristics on Fig. 4 are plotted for the following set of component values (normalized for $Z_{in}=Z_{out}=1 \Omega$, and $\omega_c=1$ Hz): $C_p=0.973$ F, $C_r=56.95$ mF, L1=1.295 H, and L2=0.647 H, so that according to equations (1-3) the angular frequencies of the passband centre and the zeros are: $\omega_c\approx 1$ rad/s, $\omega_z\approx 0.867$ rad/s and $\omega_{zz}\approx 1.227$ rad/s.

III. REALIZATION EXAMPLES

The component values can be scaled for any required central frequency and terminating impedance. For example, for a filter with $f_c=1$ GHz and $Z_{in}=Z_{out}=50 \Omega$, the component values would be: C_p=3.09 pF, C_r=0.18 pF, L1=5.15 nH, and L2=10.3 nH. The major challenge is to design a printed filter layout that accurately corresponds to the structure presented in Fig. 5, as well as to precisely fabricate the calculated values of the components and make all the resonators identical. One possible solution in the microstrip technique is proposed in Fig. 5. The overall filter is square-shaped in order to minimise the occupied space. The required inductances L1 and L2 are realised as microstrip transmission lines having а characteristic impedance between 50 Ω and 100 Ω . These approximate limits are established on a base of practical experience as a compromise between the space occupied by the transmission lines, which is lower for the lines with higher Z_c , and the conductive losses, which is lower for the transmission lines with lower Z_c . For the case of loaded (input and output) resonators the transmission lines are divided into two unequal parts by input and output ports in order to achieve the anti-parallel filter's configuration which is essential for its operation as explained in section II. The microstrip line within the resonator is meandered for minimisation of the resonator's dimensions and overall filter size, and terminated on both ends with wide microstrip patches (shaded parts on Fig.5b) that form required capacitances to ground (C_p) . Most examples presented in this paper are realized on the dielectric substrate RO3010 $(\varepsilon_r=10.2, h=0.635 \text{ mm})$. On this substrate, the microstrip patch having the capacitance C_p occupies an area of about $(\lambda_g/25)^2$ while the size of a whole filter's layout on Fig. 5a is about $(\lambda_g/6)^2$. The coupling capacitances C_r are formed between adjacent pairs of patches belonging to the neighbouring resonators. C_r 's value is determined by the gap width Sr (Fig. 5b). Gaps Z shown on Fig. 5b facilitate that the components C_p , C_r and L1 (or L2) are connected at the same node as required by configuration on Fig. 2.



Fig.5. (a) One possible layout for microstrip filter realization (b) Major dimensions for one filter's resonator

The layout in Fig. 5 introduces some parasitic elements not shown in Fig. 2. The most significant among them is the capacitance C_m formed between the two capacitive patches that belong to the same resonator, and the distributed capacitance of the microstrip transmission lines. These two types of dominant parasitic elements are included in the circuit simulator scheme in Fig. 6, which can also be used for the inclusion of dielectric and conductive losses into the calculation of the filter characteristics.



Fig.6. Circuit simulator scheme with added parasitic capacitance C_m and inductances L1 and L2 realised as microstrip transmission lines MLIN1 and MLIN2

In the next part of this paper several different designs of the capacitive coupled filters will be described and the measured results obtained from realized models will be presented.

A. Filter with Asymmetric Inductors at C Band [8]

This filter is designed for the pass band central frequency of 5.3 GHz. The scheme from Fig. 6 was used in a circuit simulator for the first-draft optimisation of the filter layout, mainly to optimise the width and the lengths of the microstrip transmission lines and to estimate the influence of the capacitance C_m , which tends to lower the filter's central frequency and to broaden the passband, as well as to take into account the components' losses. The final fine tuning of all the layout dimensions was performed by the program for EM analysis IE3D [17]. The filter was designed and realised on the dielectric substrate RO3010 at a central frequency of 5.3 GHz. Fig. 7 shows a detailed layout of an optimized filter. The dimensions [mm] indicated on Fig. 7 are: A = 2.33, $B=1.47, C=0.7, D=0.51, E=0.975, F=1.09, S_1=0.42,$ $S_2 = S_3 = 0.08$, $S_4 = 0.1$, W = 0.16, $W_0 = 0.57$, r = 0.08, R = 0.24. A photograph of the realised filter is shown on Fig. 8.



Fig.7. Layout of printed microstrip filter for f_c =5.3 GHz



Fig.8. Photograph of realized filter with f_c =5.3 GHz

Fig. 9 shows simulated and measured S_{11} and S_{21} in the frequency range from 3 to 14 GHz. The agreement between the measured results and the results from EM analysis is excellent, and even the circuit simulator gives a very good prediction for the filter characteristics. The insertion loss at f_c is 2.5 dB. The measured 3 dB and 1 dB bandwidths are 8%, and 5.9%, respectively while, the return loss in the same frequency range is better than -14 dB. The filter has a parasitic spurious passband shifted toward higher frequencies so that the insertion loss at $2f_c$ is 38 dB, while the first spurious resonance occurs at 12.9 GHz, which is 2.43 f_c . The S_{21} frequency response has two zeros of -55 dB (at 0.888 f_c) and -60 dB (at 1.226 f_c).



Fig. 9. S₂₁ and S₁₁ parameters of the filter operating at C band:
(i) measured, (ii) EM simulation results (IE3D)
(iii) circuit simulator results

B. Filter with Asymmetric Inductors at L Band [11]

The second filter is designed and realized at the central frequency of 1270 MHz on the same type of dielectric substrate (RO3010). The layout of this filter is shown on Fig. 5. It is possible to use dielectric substrates with lower ε_r , but it would increase the overall filter size. However, a selection of some thinner substrates would allow further filter size reduction. The layout of this filter cannot be obtained by simple geometry scaling of the layout from Fig. 7 because the value of the coupling capacitor C_r is inverse-proportional to the gap width S_r . Fig. 10 presents a picture of the realized filter. The filter is also square-shaped with dimensions of 16.5×16.5 mm which is a suitably small size for integration within various microwave subsystems.



Fig.10. Photograph of realized filter with f_c =1270 MHz

Fig. 11 shows the measured results of the filter's S_{21} and S_{11} frequency responses in a range from 0.5 to 3.5 GHz. The agreement between the measured results and the results from EM analysis is excellent. The realized filter has a pass-band at a central frequency of 1280 MHz (m1), which is less than 0.8% higher than designed. This difference is caused by tolerances during the filter's fabrication. The biggest influence comes from the width tolerance of the microstrip transmission lines within the filter's resonators that forms *L1* and *L2* inductances.



Fig.11. S_{21} (black) and S_{11} (gray) parameters of the L band filter: a) measured results – solid line; b) EM analysis – dotted line

The insertion loss at a central frequency is 2.4 dB. The measured 3 dB bandwidth is 7.6%, while 1 dB bandwidth is 6.3% with a return loss in the same frequency range better than 12 dB. The attenuation in the lower stop-band is around -45 dB and around -50 dB in the upper stop-band. The first spurious resonance occurs at 3.365 GHz (m5), which is 2.63 f_c , so that the insertion loss at 2 f_c is 50 dB (m4). The S_{21} frequency response has two deep minimums of -65 dB at 0.851 f_c and -80 dB at 1.238 f_c (m2 and m3).

C. Filter with Asymmetric Inductors at UHF Band [13]

The design requirements for this filter asked for a microstrip filter of an acceptable size with 3 dB bandwidth of 15%, related to the central frequency. A broader passband can be achieved by increasing the value of the coupling capacitors C_r as well as the value of the inductances ratio K=L2/L1. If scaled to a central frequency of 380 MHz the layout from

Fig. 5 would occupy too much space. Because of that as well as because of the requirements for the wider bandwidth, the design from Fig. 5 had to be modified. Also, the required value of coupling capacitors C_r couldn't be achieved only by narrowing the gaps S_r between the adjacent resonators. Instead of that, the capacitors C_r are realized as interdigital capacitors. For size minimization purposes, the value of C_p was limited and the required central frequency was achieved by increasing the value of the inductances L1 and L2. The obtained, relatively large inductance values were realized as microstrip lines that were meandered for the minimization of occupied space. The input and output microstrip lines divide the meandered microstrip line into two unequal parts in order to achieve the inductance ratio of K=3 as required for the desired width of a passband. The layout obtained according to the described process is shown in Fig. 12. The achieved printed filter is square-shaped with dimensions of 26.5×26.5 mm or $(\lambda_g/11.5) \times (\lambda_g/11.5)$ for the given dielectric substrate (RO3010).



Fig.12. Layout of printed microstrip filter for f_c =380 MHz

The microstrip layout from Fig. 12 was analyzed with the program for EM [17]. The realization of large filters inductances as heavily meandered microstrip lines as well as interdigital coupling capacitors generated a complex layout that was challenging for any EM analyzer. Because of that, the obtained results were very sensitive to the applied segmentation pattern and an edge meshing was required for an accurate characterization of the filters frequency response.



Fig.13. Photograph of realized filter with f_c =380 MHz

The filter is realized using a standard photolithography process on the dielectric substrate RO3010. The photograph of the realized filter together with a millimetre scale ruler is shown in Fig. 13. The measured results of the realized filter, as well as the EM analysis results are presented in Fig. 14. The insertion loss at a central frequency is 2.2 dB (m1), while the 3 dB passband is 19% wide. The return loss within the pass-band is - 9.5 dB or better. Two transmission zeros in proximity of the pass-band (m2, m3) show an attenuation of 49 dB@300 MHz and 67 dB@556 MHz respectively. The attenuation in the lower stop-band, from DC to the lower transmission zero, is better than 35 dB. The attenuation at $2f_c$ (m4) is 37 dB, while the lowest parasitic pass-band occurs at 980 MHz (m5).



Fig.14. S_{21} and S_{11} parameters of the UHF band filter: a) measured results – solid line; b) EM analysis – dotted line

D. Filter with Asymmetric Inductors at X Band [14]

The previous realizations have presented the filter's layout which was scalable in a wide range of microwave frequencies with an overall filter size significantly reduced even at moderate microwave frequencies of 5.3 GHz. The further scaling for higher central frequencies could lead to an impractically small filter size. Because of that a filter's layout at higher frequencies has to be adjusted so that the parasitic capacitance of transmission lines within the filters' resonators is employed to provide the required capacitances to ground (C_p) which at higher frequencies becomes very small. In addition to layer modification, a substrate with lower ε_r can be used to increase the overall filter size and make it suitable for fabrication with available technology.



Fig.15. Layout of printed microstrip filter for f_c =11.5 GHz

Fig. 15 shows a modified layout for a filter designed for a central frequency of 11.5 GHz on dielectric substrate RO4003 (ε_r =3.38, *h*=0.2 mm), while Fig. 16 shows a photograph of a realized filter. The filter has ring structure with a diameter of 5.5 mm, which is less than $\lambda_g/4$ (\approx 8 mm) for the given substrate and filter's f_c .



Fig.16. Photograph of realized filter with f_c =11.5 GHz

Fig. 17 shows results, both measured and from EM analysis, for S_{11} and S_{21} in the frequency range from 1 to 30 GHz. The agreement between the measured and predicted data is very good. The insertion loss at f_c is 2.5 dB. The measured 3 dB bandwidths are 10%, while the return loss at f_c is better than -20 dB, both measured and simulated. The insertion loss from DC to $0.87 f_c$ and from $1.28 f_c$ to $2 f_c$ is better than 35 dB. The central frequency of the filter is shifted from the designed for 1.8 % as a result of fabrication tolerances. A better agreement can be achieved if the resonator's lines and slots are fabricated with a tolerance no bigger than $\pm 5 \,\mu$ m. The S_{21} frequency response has two zeros of about $-50 \,\text{dB}$ at $0.82 f_c$ and $1.26 f_c$.

The filter has a parasitic spurious passband shifted towards higher frequencies so that the insertion loss at $2f_c$ is 38 dB, while the first spurious resonance occurs at 12.9 GHz, which is 2.43 f_c .



Fig.17. S_{21} (black) and S_{11} (gray) parameters of the X band filter: a) measured results – solid line; b) EM analysis – dotted line

E. Filter with Asymmetric Capacitors at S band [15]

In this realization example a different method for achieving the lattice filter configuration is described. Fig. 18 shows a filter structure different from those from Fig. 2. It also consists of only four different components (L, C_r , C_{pl} , and C_{p2}) that make four identical mutually coupled resonators (Q1-Q4). However, unlike the structure from Fig. 2, the resonators are unevenly loaded with grounding capacitances C_{pl} and C_{p2} , while the input and output port equally divide inductances within the input and output resonators (shaded black on Fig. 19). Like in the case of original configuration, by varying of the values: L, C_r , C_{p1} and C_{p2} it is possible to design band-pass filters at various central frequencies and with different bandwidths. This structure is more challenging for layout designing because uneven capacitances C_{p1} and C_{p2} prevent realization of symmetrical resonators (Q1-Q4), while at the same time these resonators combined together should form a structure with central symmetry. One possible layout for this filter type is shown on Fig. 19.

Compared to the original structure the new configuration has some significant advantages. With a new structure it is easier to achieve a higher value of coupling capacitance (C_r) due to the longer bordering line between adjacent resonators. This also helps when the filter has to be fabricated on substrates with lower ε_r , as well as with coarser fabrication tolerances. The filter's bandwidth is limited with the highest achievable ratio of the non-even components within the resonator (ie. L1/L2 for the first and C_{p1}/C_{p2} in the second structure) which can be significantly higher with this new configuration. The modified structure also ensures the colinearity of the input and output ports, which makes the filter suitable for integration with the surrounding components.



Fig.18. Electrical scheme of a filter with asymmetrically loaded resonators

In this case, the expression for the central angular frequency of the filter becomes:

$$\omega_{c} = \frac{1}{\sqrt{2L \frac{C_{e1}C_{e2}}{C_{e1} + C_{e2}}}},$$
(4)

where:

$$C_{e1} = C_{p1} + \frac{C_{p2}C_r}{C_{p2} + C_r}$$
(5)

$$C_{e2} = C_{p2} + \frac{C_{p1}C_r}{C_{p1} + C_r}$$
(6)

while the angular frequency of the two transmission zeros in proximity of the filter's bandpass region can be approximated as:

$$\omega_{z1} = \frac{1}{\sqrt{LC_{e1}}} \tag{7}$$

$$\omega_{z2} = \frac{1}{\sqrt{LC_{e2}}} \tag{8}$$



Fig.19. A layout of a microstrip filter corresponding to the scheme from Fig.18. Input and output resonators are shaded in black, while the two coupling resonators are shaded in gray



Fig.20. Photograph of realized filter with asymmetric capacitors



Fig. 21. S_{21} and S_{11} parameters of the S band filter: a) measured results – black line; b) EM analysis – gray line

The filter, whose layout is shown in Fig. 19 and in the photograph in Fig. 20, is designed with a program for fullwave EM analysis [17] for the central frequency of 2.55 GHz on substrate RO4003. The filter is square-shaped with overall dimensions 11.2×11.2 mm $(0.155\lambda_g \times 0.155\lambda_g)$. The gray line in Fig. 21 shows the EM analysis results for S_{21} and S_{11} parameters with the following characteristics: $IL@f_c=4dB$, $BW_{3dB}=7\%$, $BW_{1dB}=5.2\%$, $f_{z1}=2.05$ GHz $IL@(f_{z1})=65$ dB, $f_{z2}=3.16$ GHz, $IL@(f_{z2})=84$ dB, insertion loss in the lower stop-band better than 45 dB, while in the upper stop-band is better than 50 dB on frequencies below the first parasitic pass/band that occurs at 7.3 GHz. The black lane on Fig. 20 shows the following measured results for S_{21} and S_{11} : $IL@f_c=3.9$ dB, $BW_{3dB}=11.2\%$, $BW_{1dB}=6.5\%$, $f_{z1}=2.02$ GHz, $IL@(f_{zl})=60 \text{ dB}$, $f_{z2}=3.1 \text{ GHz}$, $IL@(f_{z2})=72 \text{ dB}$, the insertion loss in lower stop-band is better than 40 dB, while in the upper stop band is better than 48 dB on frequencies below the first parasitic pass/band that occurs at 7.27 GHz. The agreement between the simulated and the measured results is very good. The discrepancies between them are mostly due to the inaccurate characterization of the coupling capacitance C_r , caused by a relatively coarse segmentation during EM simulation.

F. Hybrid Filter with Asymmetric Capacitors at L band [16]

The printed filter characteristics depend significantly on the dielectric substrate used for its realization, which very often cannot be selected freely. In some cases a filter needs to be made on substrates with unfavourable characteristics like a relatively thick FR4 (h=1.6 mm) with a low and imprecisely defined ε_r (4.5 < ε_r < 5.5). The grounding printed capacitors C_p realized on this substrate could be impractically large. One possible solution in such a case could be a hybrid filter configuration with asymmetrically loaded resonators in which four lumped capacitors (one in each resonator) provide a differential portion $(C_{pc}=C_{p2}-C_{p1})$ of grounding capacitances as shown in Fig. 22. Since the lumped elements usage in filters operating at microwave frequencies is not widely used nor recommended, the four small capacitors (2.7 pF each), required for this design, are carefully selected among high quality components with a tight tolerance. This directly reduces the overall filter dimensions because of the reduced printed capacitors value(=size). Another benefit is the re-establishing of the printed resonators symmetry by equalising all the printed elements' pairs within them $(L1 = L2 = L, \text{ and } C_{p1} = C_{p2} = C_p).$



Fig.22. Electrical scheme of a filter with asymmetrically loaded resonators obtained by additional lumped capacitor C_{pc}

A two-sided layout of a hybrid filter designed for $f_c=1.27 \text{ GHz}$ ($C_p=1.8 \text{ pF}$, $C_{pc}=2.7 \text{ pF}$, $C_r=0.23 \text{ pF}$ and L=5.5 nH), is shown in Fig. 23. The pre-specified substrate was especially inadequate for the coupling capacitor's (C_r) usual realization methods. Because of that a two-layer structure is employed which allowed the C_r realisation by overlapping a rectangular portion of adjacent resonators. A pair of resonators is placed in each layer and surrounded with grounding metallization. Because of the uncertain value of the substrate's ε_r trimming patches are placed close to the printed resonators for possible fine tuning. The grounding metallization areas placed on both layers are connected together by metalized vias.



Fig.23. A layout of a two-layered printed filter corresponding to the scheme from Fig.22. Top layer contains input and output resonators (shaded in black) and horizontally shaded ground. Bottom layer contains coupling resonators (shaded in gray) and vertically shaded ground. Outlined patches are left for trimming. Coupling capacitors are realized by the overlapping of rectangular portion of adjacent resonators. The ground overlapping areas are square-shaded and contain connecting vias (not shown here)



Fig.24. Photograph of top (right) and bottom (left) side of realized hybrid filter. Lumped capacitors (C_{pc}) are marked by white pointers.



Fig. 25. S_{21} and S_{11} parameters obtained by EM analysis



Fig. 26. Measured S_{21} and S_{11} parameters of the hybrid filter

The described structure is challenging for EM analysis. More accurate models tended to overload available computation power while the simplified models provided results with limited accuracy suitable for qualitative analyses only as shown on Fig. 25. Because of this an ample number of trimming patches was included into layout for possible finetuning. However, the realized filter gave satisfactory results with quick and limited fine tuning. The active size of the realized filter is 38×38 mm $(0.3\lambda_g \times 0.3\lambda_g)$, and it can be significantly reduced (below $0.2\lambda_g \times 0.2\lambda_g$) by elimination (omission) of the trimming patches. The measured results at Fig. 26 shows the following results: $IL(f_c)=4.6 \text{ dB}$, $BW_{3dB}=7.5\%$, $BW_{1dB}=5\%$, $RL(f_c)=-12$ dB, then the insertion loss is better than 48 dB and 40 dB in the lower and upper stopband respectively. The lowest spurious passband occurs at 3.33 GHz (2.6 f_c), while IL(2 f_c)=43 dB. S_{21} transmission zeros occurs at f_{zl} =1.035 GHz (0.815 f_c) and f_{zl} =1.555 GHz $(1.225 f_c)$, both with *IL*>75 dB.

IV. CONCLUSION

In this paper an extensive overview of six filters sharing similar configuration and characteristics is presented. All filters have the lattice structure which, although hardly ever used for the realization of analogue filters, turned out to be quite suitable for filters consisting of coupled resonators. The presented examples showed that the printed filters with significant size reduction could be realized by applying exclusively the capacitive coupling between the containing resonators. All realized filters in this paper showed excellent agreement between measured and designed characteristics.

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